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(54) **Equalisation of multicarrier signals**

(57) The binary signals that are to be transmitted are associated with elementary components $(x(n,k))$ belonging to a predetermined constellation, are grouped and modulated upon N orthogonal carriers constituting OFDM symbols, preceded by a guard interval, and transmitted over a transmissive channel. The output signal from the channel is demodulated into their elementary components $(y(n,k))$, are equalized by

dividing them by respective complex values $(\hat{H}(n,k))$ constituting estimates of the channel transfer function with respect to the respective carriers before being reconverted into binary signals. The estimates of the channel transfer function $(\hat{H}(n,k))$ are filtered in the frequency domain with an equivalent "bandwidth" broader than the time duration of the guard interval.

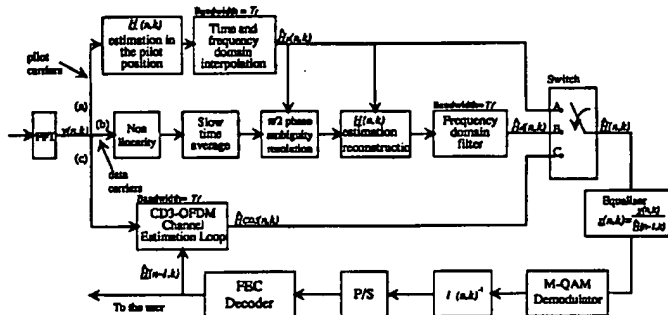


Fig. 10

Description

Background of the Invention

5 This invention is concerned with a new process and apparatus for receiving digital signals transmitted with C-OFDM (Coded Orthogonal Frequency-Division Multiplex) modulation. The invention is particularly, though not exclusively, intended for application in the reception of digital audio and video signals, and particularly of those signals that are transmitted according to the European standard ETS 300 744, "Digital broadcasting system for television, sound and data services: Framing structure, channel coding and modulation for digital terrestrial television" (denoted in the sequel, for simplicity, with the acronym DVB-T, Digital Video Broadcasting - Terrestrial).

10 An exposition of OFDM modulation can be found, for example, in S.B. Weinstein, P.M. Ebert, *Data Transmission by Frequency-Division Multiplexing using the Discrete Fourier Transform*, IEEE Transactions on Communications, COM-19(5), pp. 628-634, October 1971.

As well known, C-OFDM modulation involves the transmission of hundreds or thousands of narrow-band digital carriers that are frequency-multiplexed with a low bit-rate. These are amplitude- and phase-modulated so that their elementary components correspond to predetermined points in the amplitude-phase plane. The set of all points constitutes a "constellation". Typically, constellations of 4, 16 or 64 points are used, which are known as QPSK, 16-QAM, 64-QAM, respectively. A graphical representation of the 16-QAM constellation is shown in Figure 1, where I and Q are the in-phase and quadrature components of one carrier. The points shown are the 16 modulation states that are recognized as belonging to the constellation, corresponding to respective elementary components of the OFDM symbols (e.g., numbers 1 to 16, i.e., all the 16 possible combinations of 4 bits). In practice, the thousands of bits comprising an OFDM symbol are divided in groups or elementary components of 4 bits, and each component is associated to a respective state in the constellation. With reference to Figure 1, signal $\underline{x}(n,k)$ (where the underlining denotes a complex vector), corresponding to an elementary component of the OFDM symbol, can therefore be represented with the formula:

$$\underline{x}(n,k) = A(n,k) e^{j\Phi(n,k)}$$

where $\Phi(n,k)$ is the phase information and $A(n,k)$ is the amplitude information, and $\underline{x}(n,k)$ belongs, for instance, to one of the constellations QPSK, 16-QAM, 64-QAM. Variable n is a time-discrete variable indicating the OFDM symbol under consideration, and k is a frequency-discrete variable corresponding to the index of the C-OFDM carrier to which it refers.

In order to generate the n -th OFDM symbol for transmission, the modulator performs the discrete inverse Fourier transform (IFFT) upon N complex vectors $\underline{x}(n,k)$ ($k = 0, \dots, N-1$), some of which are null in order to make the filtering process easy. Similarly, the demodulator performs the discrete Fourier transform (FFT) upon the received signal, in order to simultaneously demodulate the N subcarriers constituting the signal. The sampling frequency f_c of the signal in the time domain (which is often, for simplicity, normalized to 1) is such that N samples are generated in the time of a useful OFDM symbol T_u , so that $T_u = N/f_c$, where N is the number of samples upon which the Fourier transform is performed.

10 In OFDM modulation, freedom from interference (i.e., orthogonality) among the carriers is insured by choosing a frequency gap among them which is equal to the useful signalling rate (symbol-rate) $1/T_u$. The symbol duration T_u is quite high, because the total bit flow for a symbol is subdivided among the several, parallel-transmitted carriers, and therefore each individual carrier occupies a very restricted portion of the spectrum. This circumstance makes OFDM modulation very robust against fading (*frequency selective fading*), as produced by echoes on the terrestrial propagation channel. In the case of a linear, frequency selective channel, the channel frequency response $\underline{H}(f)$ for the n -th symbol may be approximated, in the neighborhoods of the individual OFDM carriers, with a succession of constant, and consequently non-distorting stretches $\underline{H}(n,k)$:

$$\underline{H}(n,k) = \underline{H}(n,k) e^{j\theta}$$

50 In the diagram of Figure 2, the abscissa shows the frequency f and the ordinate shows the absolute value, or amplitude $H(n,k)$, of the transfer function of a typical transmission channel. Frequency range B is the overall bandwidth of the C-OFDM transmission.

In many cases, the $\underline{H}(n,k)$ values are substantially constant even as n varies (i.e., for successive OFDM symbols), say in the case of terrestrial television transmission to stationary receivers. However, in the case of objects moving near the antenna and in the case of mobile receivers, the diagram of Figure 2 evolves in time, sometimes quite quickly.

Moreover, in order to insure insensitivity to echoes, the C-OFDM system inserts a guard interval of duration $T_g = M/f_c$ between one symbol and the next. Such guard interval comprises M samples unused by the receivers, in

order to avoid that two contiguous symbols interfere mutually (ISI, Inter Symbol Interference). Consequently, the overall duration of a symbol, T_s , is the sum of the duration of the useful symbol T_u and of the guard interval T_g . The insertion of long guard intervals allows the system to operate in presence of even highly delayed echoes (with delays of, say, a few tens or hundreds of microseconds). However, it requires that a long symbol duration T_u is adopted, in order to maintain a good transmissive efficiency. Since the spacing between C-OFDM carriers is $1/T_u$, as T_u increases the spacing between carriers diminishes, and the number of C-OFDM carriers increases for the same bandwidth occupied by the signal.

The elementary complex signal received, $y(n,k)$, (after translation to the baseband and demodulation by means of FFT) is a replica of the transmitted signal, multiplied by the frequency response of the channel and added to a component of white, narrow-band Gaussian noise $u(n,k)$. The complex signal received is therefore:

$$y(n,k) = x(n,k) \cdot H(n,k) + u(n,k) = A(n,k) \cdot H(n,k) \cdot e^{j[\Phi(n,k) + \epsilon(n,k)]} + u(n,k)$$

Coherent demodulation involves equalization of the received signal according to the rule:

$$\hat{z}(n,k) = y(n,k) / \hat{H}(n,k) A(n,k) \cdot e^{j[\Phi(n,k)]} + v(n,k)$$

where $v(n,k) = u(n,k) / \hat{H}(n,k)$.

Such procedure requires that the receiver computes an estimate (indicated by the mark ^) of the channel frequency response $H(n,k)$. For this purpose, it is necessary to have a reliable estimate $\hat{x}(n,k)$ of the transmitted signal, and evaluate $\hat{H}(n,k)$ as:

$$\hat{H}(n,k) = \hat{y}(n,k) / \hat{x}(n,k) \quad (1)$$

Two techniques are currently suggested for estimating the transfer function of the channel for C-OFDM signals by means of formula (1):

- one technique is based on providing the C-OFDM symbol with a "comb" of modulated pilot carriers (as shown on Figure 3, where the dark cells indicate the pilot carriers) comprising vectors $\hat{x}(n,k)$ known to the receiver;
- the other technique, known as CD3-OFDM, uses the OFDM symbol received in the previous step, after correction, re-encoding and re-modulation, to get a reliable estimate $\hat{x}(n,k)$ of the transmitted sequence, and is based, for starting the equalization process, on a known symbol which is periodically transmitted (reference symbol, dark cells on Fig. 4).

In the receiver using pilot carriers (according to Figure 3), the pilot carriers are extracted from the received signal, and an estimate $\hat{H}(n,k)$ of the transfer function of the channel for all values k corresponding to a pilot carrier is derived from them by means of formula (1) (sub-sampling of $\hat{H}(n,k)$ in the frequency domain). By means of interpolations or filtering in the time and the frequency domains, or by means of a set of phase-locked loops (PLLs), an estimate of $H(n,k)$ is obtained for all values of n and of k . Sub-sampling in the time domain should be such that the channel remains quasi-stationary between two consecutive estimates; sub-sampling in the frequency domain should guarantee a number of pilot carriers of at least N/M so that all echoes with delays up to T_g are equalized (sampling theorem).

The CD3-OFDM technique yields an estimate of $H(n,k)$ for all values of k , according to the following concept (Figure 5). The binary signal issued from decoder FEC (after error correction) is re-encoded, interlaced and re-modulated by modulator M-QAM, using a procedure similar to the one performed in the transmitter, so that a reliable estimate $\hat{x}(n,k)$ of the transmitted sequence is obtained. The signal, thus reconstructed, is used to evaluate, by means of formula (1), the channel frequency response. This is then filtered in the frequency domain and the time domain to reduce the effects of noise and of wrong estimates of $\hat{x}(n,k)$.

Both techniques have been described in Italian Patent Appln. No. TO95A 258, filed on 5 April 1995 by the Applicant; in V. Mignone, A. Morello, "CD3-OFDM a novel demodulation scheme for fixed and mobile receivers", IEEE Transactions on Communications, Vol. 44, No. 9, September 1996; and in V. Mignone, A. Morello, M. Visintin, "CD3-OFDM, a new channel estimation method to improve the spectrum efficiency in digital terrestrial television systems", IBC'95 Conference, Amsterdam.

In both equalization methods as described above, the channel impulse response is assumed to be limited in time

to a duration equal to the guard interval T_g of the C-OFDM signal; consequently, a frequency-domain filter is chosen (an interpolator, for a system using pilot carriers) having a "bandwidth" T_f that is not broader than the value of the guard interval, in the former case in order to reduce aliasing due to the sub-sampling of the pilot carriers, and in both cases in order to reduce the effects of noise. It should be noted that in this case the time domain and the frequency domain are exchanged with respect to conventional filters: the channel transfer function becomes, effectively the signal to be filtered. Moreover, the following relationship holds:

$$\mathcal{F}\{\mathcal{F}\{h(t)\}\} = \mathcal{F}\{H(f)\} = h(-t)$$

where $\mathcal{F}\{\cdot\}$ denotes the Fourier transform of the argument. Accordingly, the channel impulse response $h(t)$ represents the "spectrum" of $H(f)$, leaving aside an overturn around the ordinate axis. For simplicity, overturning of the time axis will not be shown in the following description.

Such techniques therefore allow C-OFDM signals to be equalized whenever the echoes existing in the transmission channel are received with a delay with respect to the main signal which is shorter than the guard interval. If the channel impulse response (and therefore the echo delay) has a longer duration than the guard interval, the following two phenomena take place in the received signal:

a) The "tails" of the previous symbol are not exhausted within the guard interval, and the signal is affected by inter-symbol interference, ISI (Figure 6) and by inter-carrier interference (ICI).

b) The equalization process is degraded by the imperfect estimate of the channel transfer function, and the echoes having a delay longer than the guard interval appear as real interfering signals to the receiver. Indeed, according to the prior art, under both methods an estimate of the channel frequency response is evaluated by means of filtering with a "band-width" T_f equal to the guard interval: consequently, the effects of the echoes outside the guard interval upon the channel transfer function are badly evaluated by the filter.

Because of the above phenomena, both equalization methods make reception very critical whenever echoes outside the guard interval exist, and require that these are associated with only limited power levels with respect to the useful signal. i.e. power levels comparable to the power of noise or of uncorrelated interfering signals.

It should be noted that a norm has been recently passed in Europe which is directed to television transmission over terrestrial channels, denoted as DVB-T, and which uses pilot carriers, but is free from a reference symbol. Accordingly, the CD3-OFDM system as described above, including transmission of a reference symbol, would be excluded from application to the future digital terrestrial television.

Objects of the Invention

Accordingly, it is the main object of the method of reception of the invention to make the equalization system more robust even where the channel impulse response exceeds the duration of the guard interval.

Another object of the invention is to allow the equalization system CD3-OFDM to operate even where the format of the transmitted system does not include a periodically transmitted reference symbol (Figure 4), but rather a comb of pilot carriers (Figure 3), which condition is found, for instance, in the DVB-T standard.

Brief Description of the Invention

The above and other objects and advantages, such as will appear from the following disclosure, are attained by the invention by providing a method for receiving digital signals under C-OFDM modulation, having the features recited in claim 1. The invention also provides a receiving apparatus incorporating the principles of the above method, according to claim 13.

Brief Description of the Drawings

The invention will now be described in more detail in a few preferred embodiments, given by way of non-limiting examples, and with reference to the attached drawings, wherein:

Figure 1 is a schematical representation of constellation 16-QAM;

Figure 2 is a diagram of the transfer function for a typical example of a transmission channel;

Figure 3 diagrammatically shows the structure of an OFDM symbol for the DVB-T system;

5 Figure 4 diagrammatically shows the structure of an OFDM symbol for the CD3-OFDM system according to the above Italian patent appln. No. TO95A 258;

Figure 5 is a block diagram of a receiver according to the prior art (CD3-OFDM concept);

10 Figure 6 represents the effect of an echo having a delay with respect to the main signal that is longer than the guard interval;

Figure 7 is a diagram showing an example of an impulse response, with echoes longer than the guard interval T_g ;

15 Figure 8 is a diagram similar to Figure 7, in the case of an sub-sampling of the frequency response H with a distance $1/T_r$ between the samples;

Figure 9 is a diagram showing a split of the echo power into useful power C and interfering power I , with respect to the echo delay t and the parameters of the receiver;

20 Figure 10 is a block diagram of a typical embodiment of the invention to achieve lock-in of the CD3-OFDM system; and

25 Figure 11 is a graphical representation of the behavior of the phase for a signal that is the sum of two signals.

Disclosure of Preferred Embodiments

Figure 7 shows a typical example of a channel impulse response that is not limited to T_g . The estimate of the channel transfer function $\hat{H}(n,k)$ (obtained by dividing the received signal by the estimate of the transmitted signal as shown by expression (1)) is sampled, and the frequency gap between the samples is equal to the distance $1/T_r = 1/(\alpha T_u)$ between the carriers known by the receiver or estimated by it ($\alpha = 1$ in case of an estimate by the CD3-OFDM method, $\alpha = 1/3$ in case of the DVB-T standard). This means that the input signal H to the frequency-domain filter (or interpolator) exhibits a spectrum having periodical replicas repeating at distances $T_r = \alpha T_u$ (Figure 8). Accordingly, it is necessary that the bandwidth T_f of the frequency-domain filter is such that it imparts a strong attenuation in the neighbourhood of $T_r = \alpha T_u$, in order to avoid the "aliasing" phenomenon for the "main signal" (null delay), while, for echoes with delays longer than T_r , the "aliasing" phenomenon cannot be avoided. In the case that $\alpha = 1$ (CD3-OFDM receiver), the first replica of the signal appears with a very long delay and does not place any significant restraints on the bandwidth T_f , however, in case that $\alpha = 1/3$ (DVB-T norm) the filter bandwidth should be less than the theoretical limit $T_u/3$. Moreover, in order to limit the complexity of the interpolator filter, T_f should be further reduced.

40 In order to achieve a reliable estimate of the channel impulse response even in presence of echoes beyond T_g , the invention consists in using a frequency-domain filter with "bandwidth" T_f , such that all echoes that it is desired to equalize fall within its "passband" (e.g., $T_f = 2 T_g$). Obviously, it is necessary that T_f is less than T_r , so that it allows replicas of $h_d(t)$ to be filtered.

45 The following expression shows how a different filter influences the splitting of the power C_i of the i -th echo into useful power C and interfering power I , as a function of the echo delay t and of the filter bandwidth T_f (Figure 9):

50

55

$$C = \sum_i w_i C_i$$

$$I = \sum_i (1 - w_i) C_i$$

$$w_i = \begin{cases} 0 & \text{se } t < 0 & \text{(echo outside the guard interval)} \\ 1 & \text{se } 0 \leq t \leq T_g & \text{(echo within the guard interval)} \\ \left(\frac{T_u - t + T_g}{T_u} \right)^2 & \text{se } T_g < t \leq T_f & \text{(echo outside the guard interval} \\ & & \text{but equalized by the filter)} \\ 0 & \text{se } t > T_f & \text{(echo outside the guard interval} \\ & & \text{and not equalized by the filter)} \end{cases} \quad (2)$$

25 Accordingly, echoes with a delay in the range T_g to T_f contribute both as useful signals and as interfering signals. With previously known equalization systems, where $T_f = T_g$, w_i is a "window" function of duration T_g , and the power of all the echoes outside the guard interval turns out to be interfering.

Surprisingly, it turns out that by widening the bandwidth T_f of the frequency-domain filter, in allowing a better estimate of the channel transfer function to be achieved, the power received with a delay longer than the guard interval T_g is made partially useful, thus improving the overall performance of the system. Accordingly, the system achieves elimination of the second of the two degradation phenomena mentioned under (a) and (b).

Where the known symbol is present in the signal frame, the CD3-OFDM system as disclosed in the Italian patent application No. TO95A000258 mentioned above will operate in presence of echoes outside the guard interval, provided that the frequency-domain filter with "bandwidth" T_g is replaced with one having a "bandwidth" equal to the longest echo that it is desired to equalize. This involves an increase of the residual noise upon the estimate of $H(n, k)$, which increase can be compensated by installing a time-domain filter having a narrower bandwidth (for instance, for $T_f = 2 T_g$, a time-domain filter may be used which averages, in each carrier, between two successive OFDM symbols). Such concept of broadening the filter bandwidth can only be applied marginally with systems using an equalization based on pilot carriers. Indeed, with reference to the example of the DVB-T standard, using sub-sampling of the pilot carriers $\alpha = 1/3$, the interpolator filter's band-width can be widened only from T_g to a theoretical maximum of $T_u/3$, in compliance with the sampling theorem.

It can be seen from the above explanation that in the case of equalization based on pilot carriers according to the DVB-T norm, T_f is small because of the sub-sampling (i.e., the spurious replicas of the estimated transfer function are quite close to one another): consequently, due to the above restraint $T_f < T_r$ on the passband of the frequency-domain filter, the practically feasible widening of the "passband" is moderate. Much wider would be the feasible bandwidth with the CD3-OFDM equalization method according to the above Italian pat. appln. No. TO95A000258. However, in the present circumstances this method cannot be used, since it requires the presence of a reference symbol that is not contemplated in the norm as currently adopted.

In the CD3-OFDM procedure, the usefulness of the reference symbol is limited to the initial stage of the process (lock-in stage), after which equalization proceeds indefinitely, based on previous symbols, after they have been corrected, re-encoded and re-modulated. According to the invention, an improvement of the CD3-OFDM equalization method of said prior patent is provided, by which a lock-in of the CD3-OFDM procedure can be achieved even without a reference symbol, although with use of pilot carriers, as, for instance, with the DVB-T standard, even while such a sub-sampling in the frequency domain is used that makes it impossible to achieve a proper estimate of the channel transfer function in presence of echoes outside the guard interval.

If the signal frame does not contain a reference symbol, but only pilot carriers in known positions in the spectrum, the invention makes it possible to achieve lock-in of the CD3-OFDM system even in presence of echoes outside the guard interval, according to the method disclosed below. The disclosure refers to the diagram of Figure 10.

The system comprises three branches acting on the signals in reception after the FFT:

- a) Fast estimate of the channel transfer function on the pilot carriers, with an interpolator filter having a "bandwidth" $T_f \ll T_U$. This estimate is coarse, because it is limited to echoes with a delay shorter than T_f .
- b) Slow channel estimate on data carriers using a filter with wide "bandwidth" T_f which is capable of evaluating even the echoes outside the interval T_f .
- c) Fast channel estimate based on CD3, using a filter with wide "bandwidth" T_f .

It should be appreciated that branch (c) on Figure 10, as well as the bottom chain for equalization, M-QAM demodulation and FEC decoding, correspond to the CD3-OFDM scheme of Figure 5.

Stage 1: fast lock-in (switch in position A):

The transfer function $H(n,k)$ in the corresponding frequency positions is estimated in branch (a), based on the received pilot carriers, by means of formula (1), similarly to a conventional receiver based on pilot carriers. By interpolation in the time domain (or, more simply, accumulation of the pilot carriers upon four successive OFDM symbols) and in the frequency domain the estimate of the channel transfer function $\hat{H}_p(n,k)$ is obtained and is used for equalizing the signal. The interpolation in the frequency domain should be performed by a filter having a maximum bandwidth $T_f \ll T_U$ (such that it guarantees suppression of the replicas of the channel impulse response). By means of an estimate of the reliability of the signal decoded by the FEC, the "quality" of $\hat{H}_p(n,k)$ is evaluated. The estimate of the "quality" of the decoded signal can be evaluated by analyzing the rate of growth of the metrics of the Viterbi algorithm in the FEC. Alternatively, the error rate on known bytes could be measured (e.g., synchronization byte in the DVB-T standard). If the quality is sufficient to allow the CD3-OFDM system to be locked in, the switch shifts to position C and processing skips to stage 3 of the procedure directly. This can be obtained if there are no echoes with significant power and with delays longer than T_f . Otherwise, the switch shifts to position B and processing goes on to stage 2 of the lock-in procedure.

Stage 2: slow lock-in (switch in position B):

In branch (a), computation of the coarse estimate $\hat{H}_p(n,k)$ of the channel transfer function, which does not take into account the echoes outside T_f continues on the basis of the received pilot carriers. Simultaneously, a nonlinearity and a time-domain filter (on successive OFDM symbols) is applied to each of the data carriers in order to estimate the channel transfer function, leaving aside a phase ambiguity of whole multiples of $\pi/2$ radians (due to the symmetry of the M-QAM constellations to $\pi/2$ rotations). The method used for this estimate can be any of the algorithms usually suggested for the estimate of the channel (carrier recovery) in digital, coherent-demodulation systems. For instance, the nonlinearity could be raising the received signal to the fourth power (an algorithm described in D. Efstathiou, P. Fines, A. H. Aghvami, "Preamble-less Non-Decision-Aided (NDA) Techniques for 16-QAM Carrier Phase Recovery and Gain Error Correction, for Burst Transmissions", proceedings of the GLOBECOM '93 Conference, 29 November - 2 December 1993). The time-domain filter can be implemented by an average operator over n successive OFDM symbols; in that case the estimate $\hat{H}_a(n,k)$ will be obtained as:

$$|\hat{H}_a(n,k)| = \sqrt[4]{\frac{\sum_{q=0}^n y^4(q,k)}{E[x^4]}}$$

$$\arg(\hat{H}_a(n,k)) = \frac{\arg(y^4(q,k)) + \pi}{4} + m\frac{\pi}{2} \quad \forall m \in \{0,1,2,3\}$$

where $E[\cdot]$ denotes the average value.

The invention resides in the method for resolving the $m\pi/2$ phase ambiguity in the estimate $\hat{H}_a(n,k)$.

Figure 11 graphically shows the signal resulting from the sum of a main signal and an echo, for two different echo amplitudes. The phase of the echo vector can take any value depending on the associated delay, as shown by the circumference in the drawing. It can be seen that the maximum phase difference of the resulting vector from the main signal increases as the amplitude of the echo becomes similar to the one of the main signal.

As mentioned above, the estimate algorithm based on pilot carriers in branch (a) allows the phase of the channel transfer function to be evaluated, for echoes having a delay shorter than the band T_f of the frequency interpolator filter. This means that, when echoes with a delay longer than T_f are present, the phase error in the estimate $\hat{H}_p(n,k)$ will be the larger, the greater is the amplitude of such echoes. However, if such error turns out to be less than $\pi/4$ in absolute value, then the phase information contained in $\hat{H}_p(n,k)$ can be used to resolve the phase ambiguity in $\hat{H}_d(n,k)$, by associating to it the phase closer to the one of $\hat{H}_p(n,k)$. From Figure 9 it can be seen that, as the echo delay increases, there is a reduction of the useful power contribution and an increase of the contribution from the interfering power associated with the echo. Even if the equalization process is assumed to be perfect, the systems under consideration can only operate correctly with C/I ratios larger than a limit value $(C/I)_L$, which depends mainly on the chosen constellation and on the adopted type of FEC coding. Consequently, if the condition $C/I > (C/I)_L$ is imposed in formula (1), it follows that the equalization system should be capable of correctly operating with echoes outside the guard interval T_g , having an amplitude decreasing as the the associated delay increases. Therefore, if bandwidth T_f of the interpolator filter in branch (a) yields a good estimate of high-power echoes within and closely outside the guard interval up to T_f , the method of the invention provides correct resolution of the phase ambiguity for echoes having a delay longer than T_f up to T_f' and a reduced amplitude, which generate phase rotations that are usually smaller than $\pi/4$ in absolute value.

Therefore, according to the invention, the estimate $\hat{H}_d(n,k)$ of the channel transfer function is given by:

$$|\hat{H}_d(n,k)| = |\hat{H}_a(n,k)|$$

$$\arg(\hat{H}_d(n,k)) = \arg(\hat{H}_a(n,k)) \text{ with } m = m_0$$

where m_0 is the value of m such that the phase difference $[\arg(\hat{H}_a(n,k)) - \arg(\hat{H}_p(n,k))]$ is at a minimum (resolution of the $\pi/2$ phase ambiguity).

Subsequently, the channel estimate as derived from the data carriers is integrated with the estimate evaluated from the pilot carriers, and the result is filtered in the frequency domain using a "wide" filter having "bandwidth" T_f' . It should be noted that, in the hardware implementation, the frequency-domain filters in branches (a), (b) e (c) could be a single physical filter, where the coefficients and the bandwidth are changed depending on the stage ((a), (b), o (c)) of the algorithm.

A second method for resolving the phase ambiguity is based on the following consideration. The phase jumps between successive samples of the transfer function $H(n,k)$ depend on the amplitude of the echo and on its delay (Figure 11). The shorter is the echo delay, the smaller is the phase variation from one frequency sample to the next in $H(n,k)$; the smaller the echo amplitude, the smaller is the phase variation from one frequency sample to the next in $H(n,k)$. Therefore, since an estimate of the transfer function $H(n,k)$ is available to the receiver for values k corresponding to the pilot carriers (after filtering in the time domain), such estimate can be used to resolve the phase ambiguity of the transfer function corresponding to the data carriers. The invention provides for associating $\hat{H}_d(n,k)$ with the phase closest to the phase of the pilot carrier that is adjacent in the frequency domain. This algorithm is sub-optimal with respect to the previous one, because powerful echoes, even inside the guard interval, could give rise to large phase jumps from one carrier to the next, and, consequently, to a wrong resolution of the phase ambiguity.

Variations in the above described methods of resolving the phase ambiguity ($m\pi/2$) may be made, which will in each case be based on phase information from the pilot carriers and from the adjacent data carriers for which the phase ambiguity has been previously resolved, all within the scope of the invention. For instance, another approach could be to search for the minimum phase difference with respect to a data carrier of adjacent frequency, for which the phase ambiguity has already been resolved in the same way, or with another of the above described methods.

In stage 2 (switch in position B), the quality of $\hat{H}_d(n,k)$ is again evaluated by estimating the reliability of the signal as decoded by the FEC, until such quality becomes sufficient to insure convergence of the CD3-OFDM algorithm. Alternatively, a fixed lock-in time could be identified (the number of OFDM symbols over which the time average should be computed), on the basis of a laboratory test using a variety of echo patterns. At this point, the switch is shifted to position C and processing proceeds to stage 3.

Stage 3. channel estimate by the CD3-OFDM method (switch in position C):

This stage coincides with the end of the lock-in stage in the estimator of $H(n,k)$, when the CD3-OFDM equalization algorithm can operate correctly with a filter that has been widened to bandwidth T_f' . This stage continues for the remaining duration of the reception, until, for any reason, the loop should become unlocked.

The result of the estimate of $H(n,k)$ in stage 2, with either of the methods suggested for resolving the phase ambiguity, is all the more accurate as the number of averaged symbols Q is larger, usually a few tens; this, of course,

involves a slowdown in tracking the channel transfer function, although, in the preferred embodiment of the receiver (Fig. 10), this only happens in the start-up stage of the receiver. Indeed, the CD3-OFDM channel estimate procedure, once triggered, is able to perform the estimate symbol by symbol, and to quickly track all changes in the channel characteristics.

5 Stage 2 of the method according to the invention may also continue for the entire duration of the reception, without using the CD3-OFDM channel estimate algorithm, and therefore without a transition to stage 3. In that case, the slow time-domain filter in branch (b) drastically slows down the ability of the system to track changes in the propagation channel (e.g., due to Doppler effect) for the entire duration of the reception. Accordingly, though applying the invention to the equalization system based on the CD3-OFDM system requires a quasi-stationary channel only for a brief lock-in
10 interval, using the invention with the equalization system based on the pilot carriers alone does not provide a sufficient adjustment speed in presence of channel variations and echoes outside the guard interval.

Claims

- 15 1. A method for receiving digital signals in C-OFDM-type multicarrier modulation, wherein binary signals for transmission are associated to elementary components ($x(n,k)$) belonging to a predetermined constellation, groups of N elementary components, each preceded by a guard interval, are modulated on respective carriers of OFDM symbols, a transmission in the transmission channel is performed, the output signals from the channel are demodulated by reconstructing the elementary vectors ($\underline{x}(n,k)$), which are then equalized by dividing them by respective complex
20 values ($\underline{H}(n,k)$) constituting estimates of the channel transfer function with respect to the respective carriers before being reconverted into binary signals, characterized in that the estimates of the channel transfer function ($\underline{H}(n,k)$) are filtered (or interpolated) in the frequency domain with an equivalent "band-width" (T_f) larger than the time duration of the guard interval (T_g) but smaller than the time gap between the periodical repetitions of the channel impulse response due to sub-sampling in the frequency domain ($T_r = \alpha T_u$).
- 25 2. Method for receiving digital signals according to claim 1, wherein the transmitted signal is of the type including a reference symbol transmitted at every predetermined group of OFDM symbols, characterized in that the estimate of the channel transfer function ($\underline{H}(n,k)$) for a current received symbol ($\underline{x}(n,k)$) is evaluated by dividing the demodulated components of a previously received OFDM symbol ($\underline{x}(n-1,k)$) by the respective estimates of the same previously transmitted symbol ($\underline{x}(n-1,k)$), as obtained by the receiver after equalization, FEC correction, re-encoding and re-mapping, and in that the initial estimate of the channel transfer function ($\underline{H}(n,k)$) is computed by dividing the demodulated components of a predetermined OFDM reference symbol by the respective components of the
30 same transmitted symbol as stored in the receiver.
- 35 3. Method for receiving digital signals according to claim 1, wherein the transmitted signal is of the type including pilot carriers without a reference symbol, characterized in that an initial estimate of the channel transfer function ($\underline{H}_p(n,k)$) is obtained in a first stage by extracting the pilot carriers from the signal, deriving the channel response for each pilot carrier and interpolating over the entire channel by a filtering in the frequency domain with a "band-width" (T_f) that is not greater than the duration of repetition interval ($T_r = \alpha T_u$) of the sub-sampled channel impulse response; and in that the estimates of the channel transfer function ($\underline{H}(n,k)$), in a second stage, are derived for each symbol by dividing the elementary components of a symbol received before the symbol under consideration by the respective estimates of the elementary components of said previously transmitted symbol, as reconstructed in the receiver after equalization, FEC correction, re-encoding and re-mapping, whereby the filtering of ($\underline{H}(n,k)$) in the frequency domain being performed with a "bandwidth" (T_f') greater than the time duration of the guard interval
40 (T_g) and equal to the maximum delay of the echoes that it is desired to equalize.
- 45 4. Method for receiving digital signals according to claim 3, wherein the received, demodulated and equalized signal is decoded by FEC decoding, characterized in that the reliability of said initial estimate of the channel transfer function ($\underline{H}_p(n,k)$) is evaluated by analyzing the rate of growth of the metrics in the Viterbi algorithm of the decoding.
- 50 5. Method for receiving digital signals according to claim 3, characterized in that the reliability of said initial estimate of the channel transfer function ($\underline{H}_p(n,k)$) is evaluated by measuring the error rate in predetermined known bytes that are contained in the transmitted signal.
- 55 6. Method for receiving digital signals according to claim 4 or 5, characterized in that, if during said first stage the reliability of the initial estimate of the channel transfer function ($\underline{H}_p(n,k)$) is worse than a predetermined value, a non-linearity and a time-domain filter are applied to the data carriers in a parallel processing in order to improve the estimate of the channel transfer function, with a phase ambiguity of multiples of $\pi/2$ radians, and the ambiguity is

resolved by assuming as a valid phase the one for which the phase difference from the initial estimate of the channel transfer function ($\hat{H}_p(n,k)$) is a minimum.

- 5 7. Method for receiving digital signals according to claim 4 or 5, characterized in that, if during said first stage the reliability of the initial estimate of the channel transfer function ($\hat{H}_p(n,k)$) is worse than a predetermined value, a non-linearity and a time-domain filter are applied to the data carriers in a parallel processing in order to improve the estimate of the channel transfer function, with a phase ambiguity of multiples of $\pi/2$ radians, and the ambiguity is resolved by assuming as a valid phase the one whose difference from the phase of the pilot carrier that is closer in frequency is a minimum.
- 10 8. Method for receiving digital signals according to claim 4 or 5, characterized in that, if during said first stage the reliability of the initial estimate of the channel transfer function ($\hat{H}_p(n,k)$) is worse than a predetermined value, a non-linearity and a time-domain filter are applied to the data carriers in a parallel processing in order to improve the estimate of channel transfer function, with a phase ambiguity of multiples of $\pi/2$ radians, and the ambiguity is resolved by assuming as a valid phase the one having minimum difference from the phase of a pilot carrier or data carrier, which is adjacent in frequency, of which the phase ambiguity has already been resolved.
- 15 9. Method for receiving digital signals according to claim 6 or 7 or 8, characterized in that said non-linearity is a raising of the received signal to the fourth power.
- 20 10. Method for receiving digital signals according to claim 6, 7, 8 or 9, characterized in that the time-domain filtering is stopped after a predetermined number of OFDM symbols and said second stage is started.
- 25 11. Apparatus for receiving digital signals, having means for receiving and demodulating signals transmitted over a transmissive channel as a sequence of OFDM multi-carrier symbols, each preceded by a guard interval, and each comprising a plurality of elementary components ($x(n,k)$) belonging to a predetermined constellation and suitably modulated onto OFDM carriers, and means for equalizing the demodulated elementary signals ($y(n,k)$) by dividing them by respective complex values ($\hat{H}(n,k)$) constituting estimates of the channel transfer function with respect to the respective carriers before being reconverted into binary signals, characterized in that it further comprises a frequency-domain filter for filtering the estimates of the channel transfer function ($\hat{H}(n,k)$) with an equivalent "bandwidth" (T_f) that is greater than the time duration of the guard interval (T_g) but smaller than the time distance between the periodical repetitions of the channel impulse response due to sub-sampling in the frequency domain ($T_r = \alpha T_u$).
- 30 12. Apparatus for receiving digital signals according to claim 11, wherein the transmitted signal is of the type including a reference symbol transmitted at every predetermined group of OFDM symbols, characterized in that it comprises means for generating an estimate of the channel transfer function ($\hat{H}(n,k)$) by dividing the demodulated elementary components of a previously received OFDM symbol by the respective estimates of the same previously transmitted symbol, as obtained after equalization, FEC correction, re-encoding and re-mapping, and in that the initial estimate of the channel transfer function ($\hat{H}(n,k)$) is obtained by dividing the demodulated components of a predetermined OFDM reference symbol by the respective components of the same transmitted symbol as stored in the receiver.
- 35 13. Apparatus for receiving digital signals according to claim 11, wherein the transmitted signal is of the type including pilot carriers without a reference symbol, characterized in that it comprises first means for obtaining an initial estimate of the channel transfer function ($\hat{H}_p(n,k)$) by extracting the pilot carrier frequencies from the signal, deriving the channel response for each pilot carrier and interpolating over the entire channel by frequency-domain filtering with a "bandwidth" (T_f) that is not greater than the duration of repetition interval ($T_r = \alpha T_u$) of the sub-sampled channel impulse response; second means for deriving estimates of the channel transfer function ($\hat{H}(n,k)$) for each successive symbol by dividing the elementary components of a symbol received before the symbol under consideration by the respective estimated components of said previously transmitted symbol, as reconstructed in the receiver after equalization, FEC correction, re-encoding and re-mapping, the frequency-domain filtering of $\hat{H}(n,k)$ being performed with a "band-width" (T_f') greater than the time duration of the guard interval (T_g) and equal to the maximum delay of the echoes that it is desired to equalize; and switching means for switching the operation
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55 of the apparatus from the first means to the second means.
14. Apparatus for receiving digital signals according to claim 13, characterized in that it further comprises third means for applying a non-linearity and a time-domain filtering to the data carriers in order to improve the estimate of chan-

nel transfer function, with a phase ambiguity of multiples of $\pi/2$ radians, and means for resolving the phase ambiguity by assuming as a valid phase the one for which the phase difference from the initial estimate of the channel transfer function ($\hat{H}_p(n,k)$) is a minimum, and in that said switching means, if the reliability of the initial estimate of the channel transfer function ($\hat{H}_p(n,k)$) in said first stage turns out to be worse than a predetermined value, switch the operation of the apparatus from the first to the third means before switching to the second means.

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15. Apparatus for receiving digital signals according to claim 13, characterized in that it further comprises third means for applying a non-linearity and a time-domain filtering to the data carriers in order to improve the estimate of the channel transfer function ($\hat{H}(n,k)$), with a phase ambiguity of multiples of $\pi/2$ radians, and means for resolving the phase ambiguity by assuming as a valid phase the one for which the phase difference from the phase of the pilot carrier closest in frequency is a minimum, and in that said switching means, if said reliability of the initial estimate of the channel transfer function ($\hat{H}_p(n,k)$) in said first stage is worse than a predetermined value, switch the operation of the apparatus from the first to the third means before switching to the second means.

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16. Apparatus for receiving digital signals according to claim 13, characterized in that it further comprises third means for applying a non-linearity and a time-domain filtering to the data carriers in order to improve the estimate of the channel transfer function ($\hat{H}(n,k)$), with a phase ambiguity of multiples of $\pi/2$ radians, and means for resolving the phase ambiguity by assuming as a valid phase the one having minimum difference from the phase of pilot carrier or a data carrier, which is adjacent in frequency, of which the phase ambiguity has already been resolved, and in that said switching means, if said reliability of the initial estimate of the channel transfer function ($\hat{H}_p(n,k)$) in said first stage is worse than a predetermined value, switch the operation of the apparatus from the first to the third means before switching to the second means.

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17. Apparatus for receiving digital signals according to any of claims 14, 15 or 16, characterized in that said means for applying a nonlinearity are means for raising the received signal to the fourth power.

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18. Apparatus for receiving digital signals according to any of claims 14 to 17, characterized in that said switching means switch from said third stage to said second stage after reception of a predetermined number of OFDM symbols.

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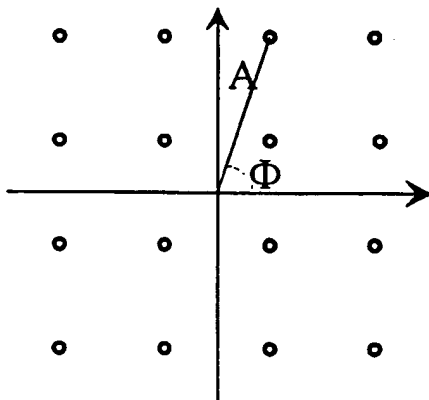


Fig. 1

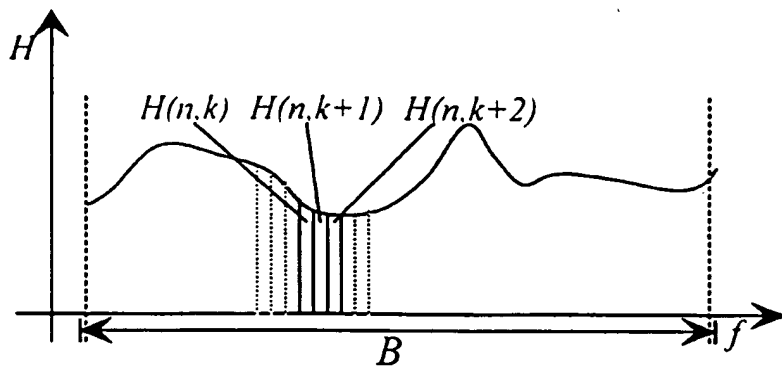


Fig. 2

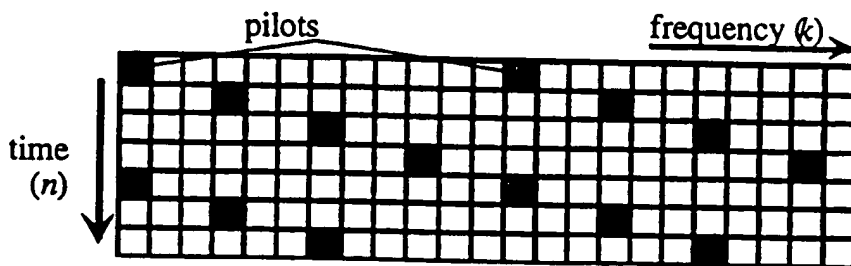


Fig. 3

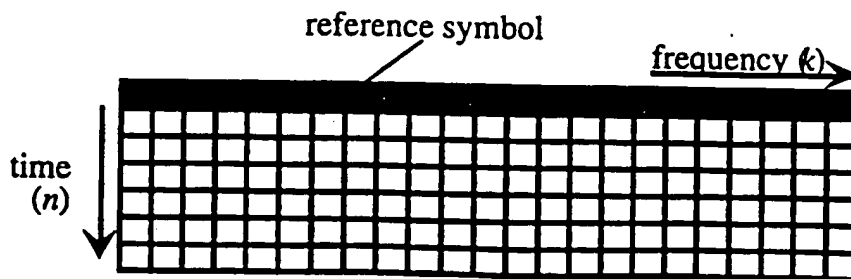


Fig. 4

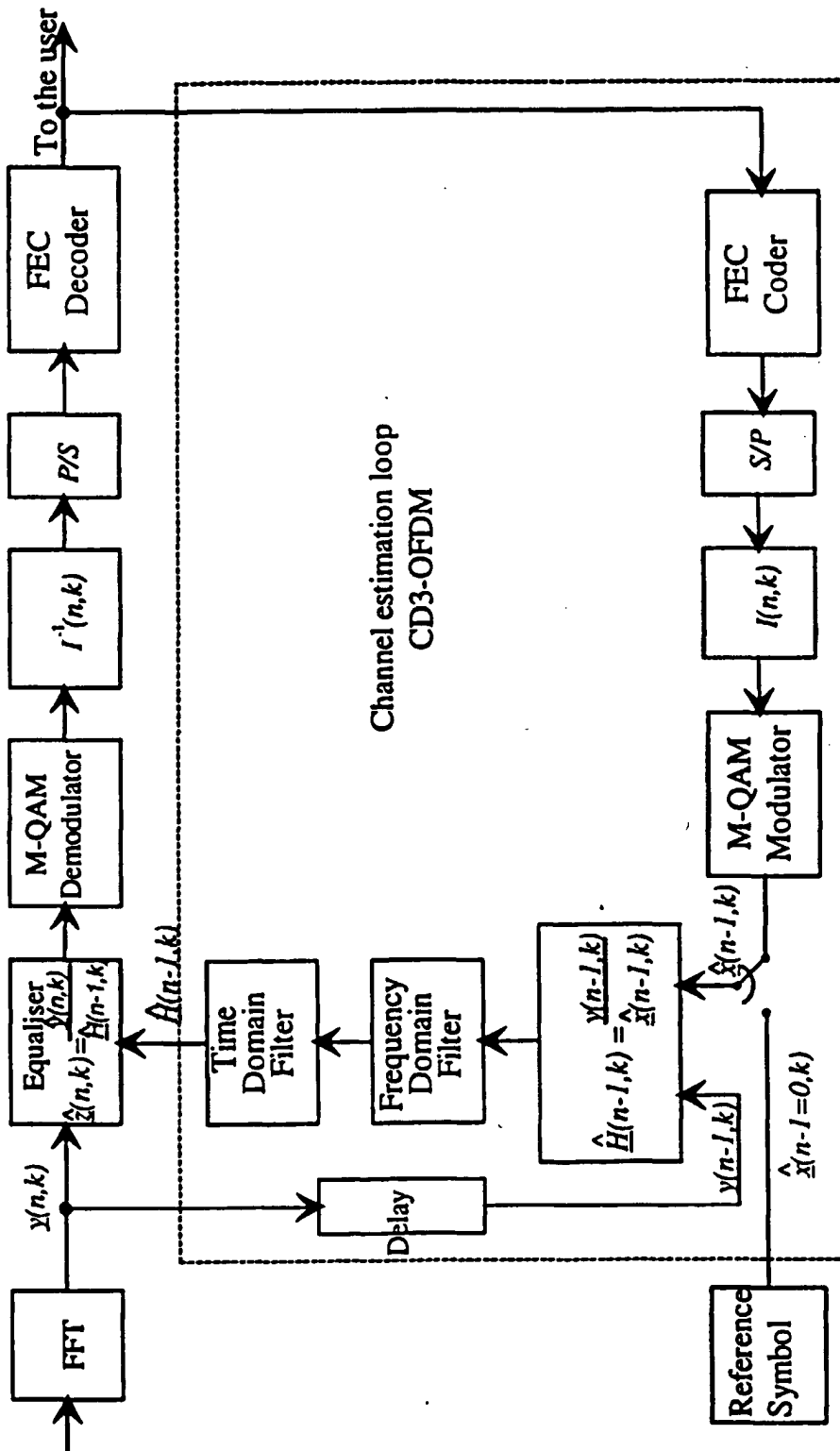


Fig. 5

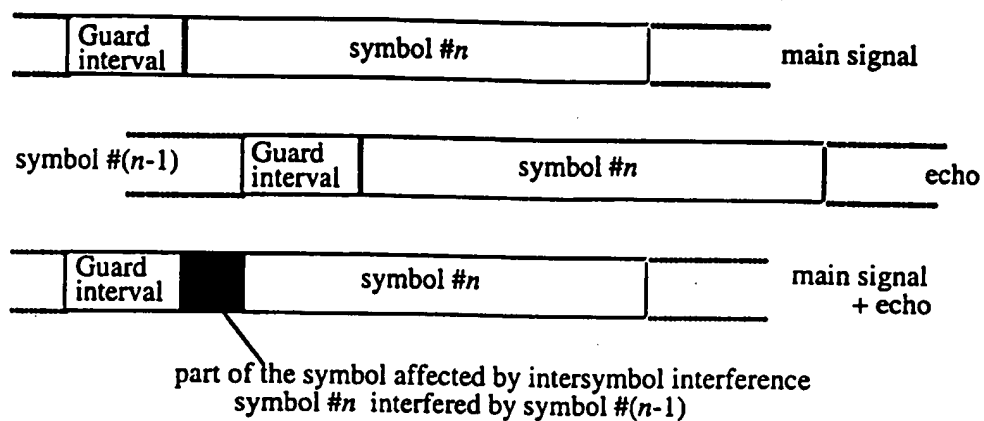


Fig. 6

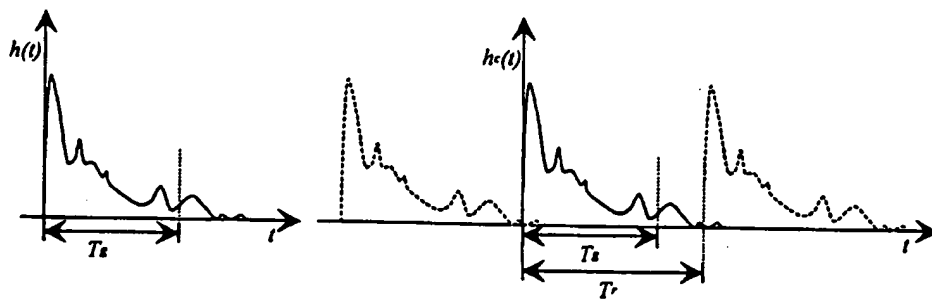


Fig. 7

Fig. 8

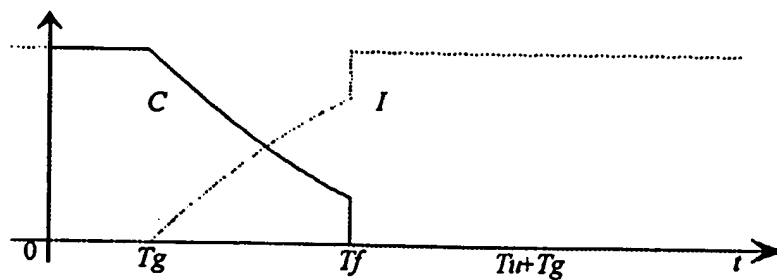


Fig. 9

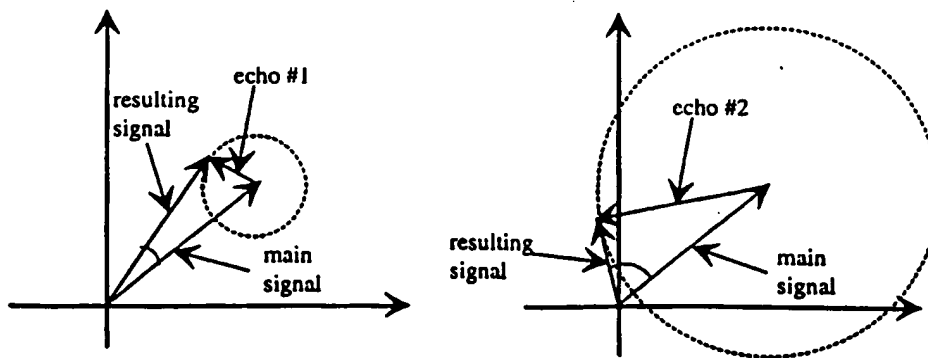


Fig. 11